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A 0.3- μ m CMOS 8-Gb/s 4-PAM Serial Link Transceiver

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Abstract

An 8-Gb/s 0.3- μ m CMOS transceiver uses multilevel signaling (4-PAM) and transmit pre-shaping in combination with receive equalization to reduce ISI due to channel low-pass effects. High on-chip frequencies are avoided by multiplexing and demultiplexing the data directly at the pads. Timing recovery takes advantage of a novel frequency acquisition scheme and a linear PLL with a loop bandwidth >30MHz, phase margin >48° and capture range of 20MHz without a frequency acquisition aid. The transmitted 8-Gbps data is successfully detected by the receiver after a 10-m coaxial cable. The 2mm x 2mm chip consumes 1.1W at 8Gbps with a 3-V supply.

Introduction

As the demand for higher data-rate communication increases, low-cost, high-speed serial links using copper cables become more attractive for distances of 1 to 10 meters[1],[2]. For multi-Gbps applications, the data rate is limited by the cable skin-effect loss and the process technology. The 10-m coaxial cable (PE-142LL) used in this work has a -3dB bandwidth of 1.2GHz. This design differs from existing Gbps links [1],[2] in its use of a receiver equalizer in combination with a transmitter filter to compensate for the cable characteristics. High on-chip frequencies are avoided by multiplexing and demultiplexing the data directly at the pads. To reduce the symbol rate, a 4-level pulse amplitude modulation (4-PAM) is used. A new proportional phase detector for data recovery is proposed which does not suffer from the stability and bandwidth limitations of traditional bang-bang loops. A novel frequency acquisition architecture is designed to enable the receive PLL to lock to the input stream under all process variations.

System Architecture

Implementing optimal detection methods for multi-Gbps rates demands high complexity and large area [3]. Instead, square pulses, which can be generated and detected with modest complexity, are used as the basis communication symbols [4]. At rates well above the channel bandwidth, however, square pulses result in severe intersymbol interference (ISI) which reduces the data-eye openings. For a given data rate, the 4-PAM scheme reduces the symbol-rate to half compared to a conventional 2-PAM system. This symbol rate reduction lowers not only the ISI in the channel, but also the maximum required on-chip clock frequency.

To invert the channel, a 1-tap equalizer at the receiver and a 2-tap pre-emphasis filter at the transmitter are used. The

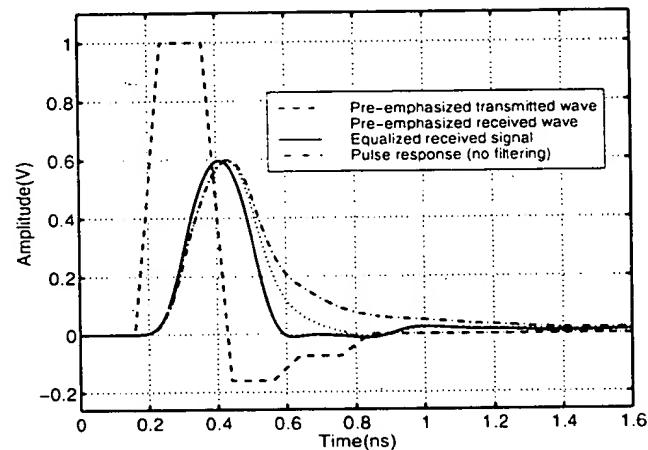


Fig. 1 Pulse shape with and without filtering

effects of these filters for a 0.2-ns pulse (5Gsym/s) at the near and far end of the 10-m channel are shown in Fig. 1. The unfiltered pulse response remains at a large value 0.2ns after its peak (next symbol sample point), while the pre-shaped equalized signal is zero at that point. The filter tap weights can be programmed for different channels.

The on-chip frequency requirement is further reduced to 1/5 the symbol rate (1/10 bit rate) by performing 5:1 multiplexing and 1:5 demultiplexing directly at the chip pads, allowing 5 symbols to be transmitted every clock cycle. The 5 symbols correspond to 10 bits that include 4 data symbols and 1 symbol for line coding. In this design, coding is performed on chip to provide enough transitions for clock recovery.

Circuit Implementation

The block diagram of the complete transceiver chip is depicted in Fig. 2. The transmitter, comprising 5 identical drivers, uses different clock phases from a 5-stage differential ring oscillator (TX-VCO) to multiplex the data stream onto the 50- Ω line. The detailed transmitter design is described in [5].

The receiver performs 1:5 demultiplexing by sampling the signal with 5 out of 10 clock phases from a 5-stage differential ring oscillator (RX-VCO). The 5 additional alternate clock phases provide required samples for the input equalizer with half-symbol-spaced tap spacings, and allow 2x oversampling to recover timing. Samples are next filtered by the 1-tap input equalizer which differentially subtracts the weighted value of the former sample from the present sample. This operation is done by current summing two differential values with opposite polarity (Fig. 3). The equalizer further improves the data eye area (height x width) up to 40% by sharpening the signal transitions. The ADC samplers are similar to those used in [7].

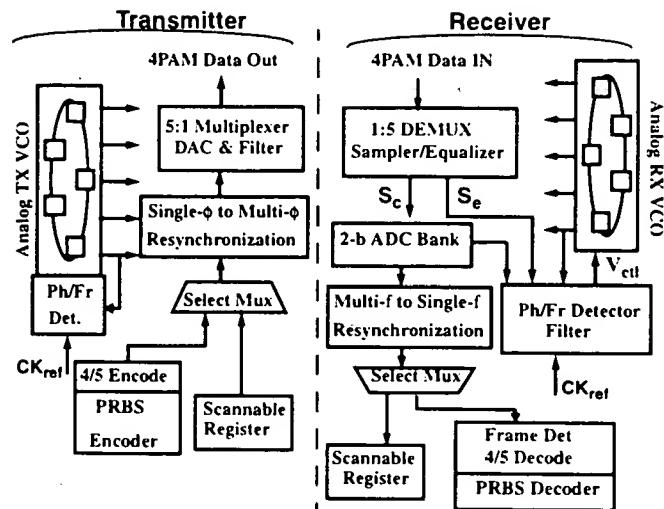


Fig. 2 Transceiver general architecture

When the receive PLL is locked properly to the input data, half of the 10 analog equalizer outputs represent sampled values at the center of the symbols (S_c), and half are samples at the data transitions (S_e). The S_c samples are digitized by 2-bit flash ADCs and result in the received data bits that are next resynchronized to a global clock. The S_e samples are buffered by linear amplifiers and used as part of a linear phase detector for timing recovery. The samples are pipelined properly to match the delays of the different signal paths corresponding to S_c and S_e .

Timing recovery uses data transitions to detect the phase between the data and the sampling clocks. In a differential 4-PAM stream, there are 3 distinct transition types (Fig. 4). Of these 3 types, only *type1* makes a transition to the same magnitude but opposite polarity, which results in a zero crossing that occurs exactly at the mid-point between two

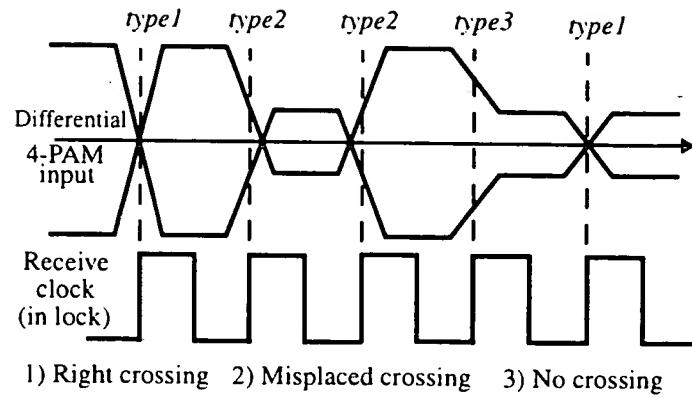


Fig. 4 Three types of transitions in a 4-PAM symbol stream

symbols and which therefore can be used for clock recovery. The two other types are ignored as they convey wrong phase information. In every cycle (5 symbols), one *type1* transition is guaranteed by the transmitter's 4/5sym encoder.

Traditional linear data PLLs offer good loop stability and bandwidth, but most suffer from a systematic phase offset. Sampling transitions by the same mechanism as the data reduces the systematic phase offset in data recovery. However, conventional sampling digital loops use bang-bang control, resulting in limited bandwidth and stability [6]. Fig. 5a shows the block diagram of the data Phase/Freq. detector used here to overcome these problems. When the data loop is in lock, the analog edge samples (S_e) at *type1* transitions are zero, resulting in zero sum current at the *charge-pump1* input ($V_p=0$). Thus, ideally no charge is pumped into the VCO control line when in lock and therefore there is no phase error due to control voltage ripple. When not in lock, the *type1* edge samples occur before (*Early*) or after (*Late*) the zero crossings, resulting in non-zero values for S_e samples. Based on the previous and next symbol values (2 bit data from ADCs), the decision logic of each stage adds the analog S_e values of *type1* with correct polarity to the control voltage of the loop and ignores the other two types of transitions by turning off the current of that stage (Fig. 5a). The chosen scheme for decision logic introduces a problem discussed in the next section. As the S_e values are proportional to the loop phase error, the correction on the loop control voltage is also proportional to the phase error. The phase detector is therefore linear. Thus this PLL combines the advantages of both a linear and a sampling loop. The loop gain, and consequently the bandwidth, increases with the number of useful (*type1*) transitions per cycle. Assuming a random data sequence with a 4/5sym encoder, the average *type1* transition density is two per cycle, which results in a loop bandwidth of 30MHz ($BW/f_{ref} > 0.06$) and phase margin $>48^\circ$ with no systematic phase offset.

As the phase detector has a limited frequency capture range, a frequency acquisition aid is employed to help acquire lock to a reference clock at start-up (Fig 5a). The circuit, shown in Fig. 5b, uses cycle-slipping information when the Rx-VCO frequency is different from that of the incoming data. During cycle-slipping, sweeping of the clock phase causes the phase

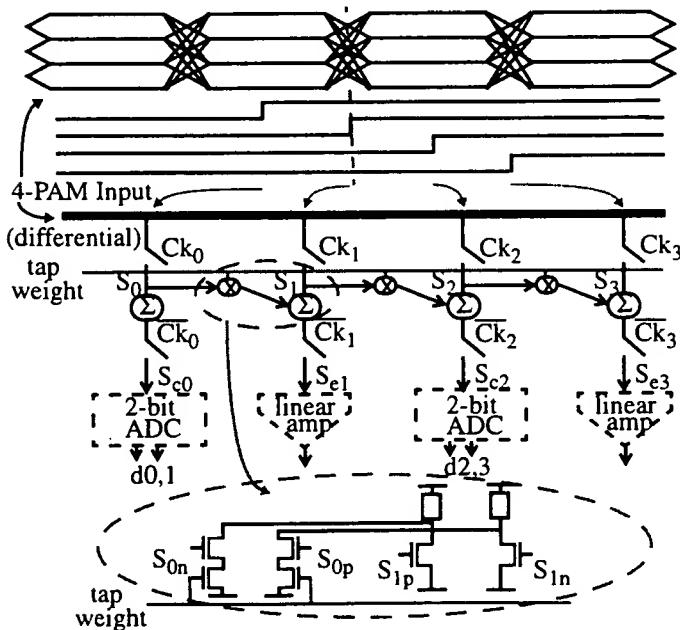


Fig. 3 1:5 Demux samplers and equalizers

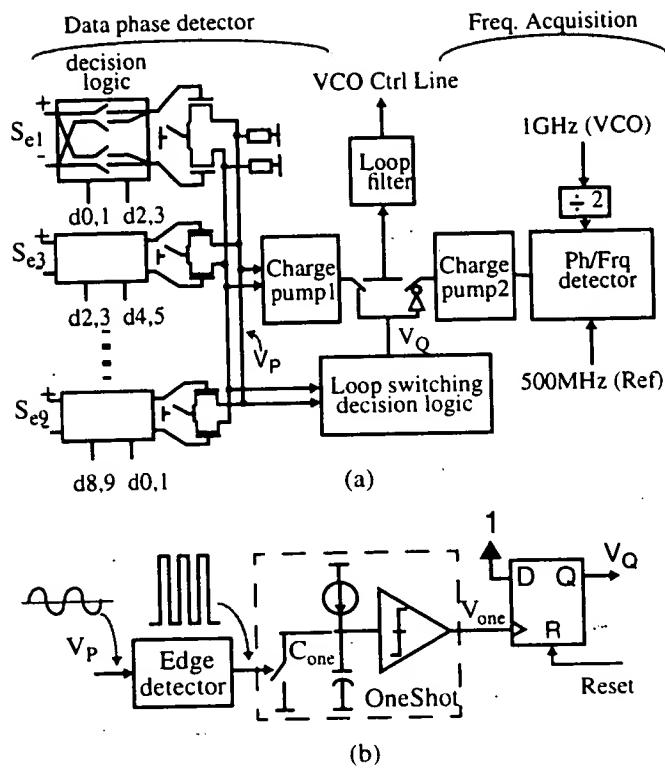


Fig. 5 a) Phase/Freq. detector architecture b) Loop switching decision logic

detector output (V_P) to oscillate between *Early* and *Late* signals. The frequency of this oscillation (sweep speed) is equal to the frequency difference between the receive clocks and the incoming data. If there is a considerable frequency difference, the oscillations at V_P cause the edge-detector (Fig. 5b) to produce pulses that continuously set the one-shot circuit output (V_{one}) to one. Once the VCO frequency is close enough to the incoming data frequency (within the data PLL capture range), the pulse rate of the edge-detector decreases such that C_{one} can charge high enough to switch V_{one} to zero. At the falling edge of V_{one} , V_Q , which is reset to zero at start-up, is asserted and hands the loop control to the data phase detector. The edge detector is designed to have hysteresis (Fig. 6), using positive feedback in its first stage amplifier. Thus, it reacts only to oscillation amplitudes larger than a certain threshold level, which helps prevent erroneous transitions due to noise.

A 2^7 -1 PRBS encoder and decoder, as well as a scannable transmit/receive data register, are provided on chip for BER testing. The 5/4sym decoder removes the extra line-code symbol (Fig. 2).

Measurements

The transmitter achieves a symbol rate of 5Gsym/s (10Gb/s) with an eye opening of 200mV and 90ps, and 4Gsym/s (8Gbps) with an eye opening of 350mV and 110ps over 10 meters of coaxial cable, using pre-emphasis (Fig. 6). Symbols without pre-emphasis after the 10-m cable show an eye opening with 60-mV height and 50-ps width at 4Gsym/s. The transmitter output has an adjustable amplitude with a

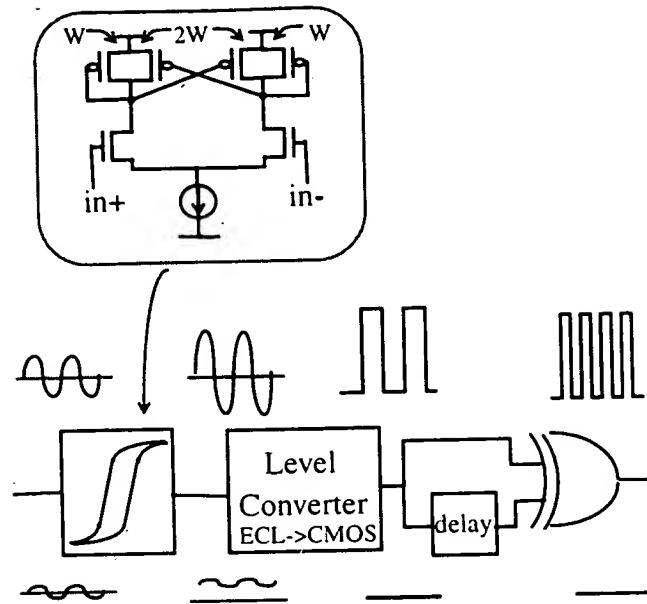


Fig. 6 Edge detector circuit

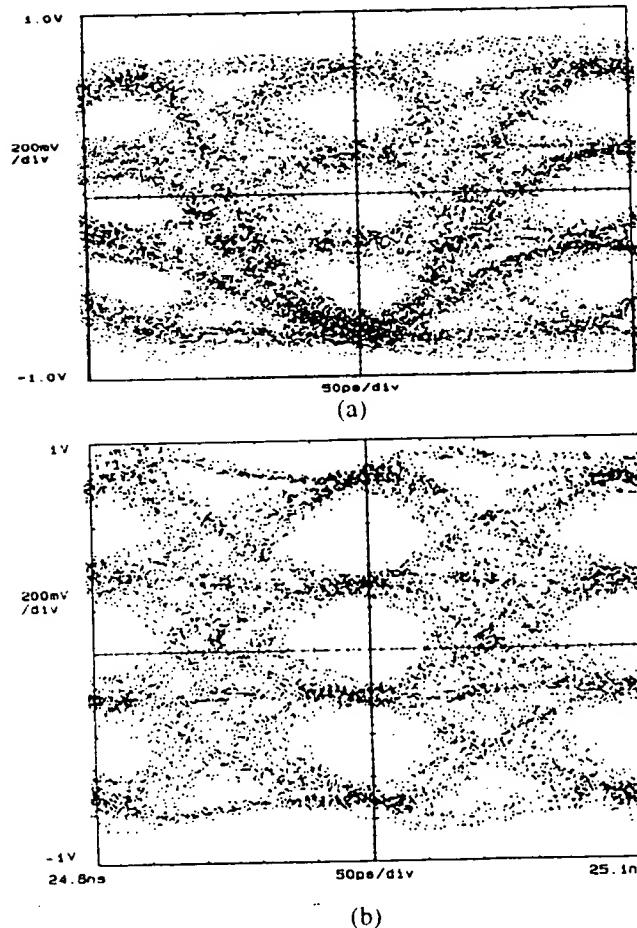


Fig. 6 Differential data-eye over 10-meter cable with pre-emphasis
a) at 10Gbps, and b) at 8Gbps

maximum of 1.2V, and a jitter of 11ps (p-p) and 2ps (rms).

The receiver successfully detects an 8-Gbps 4-PAM data stream after 10 meters with a 3-V supply. Receiver equali-

tion helps reduce the transmitter pre-emphasis for the 10-m cable, effectively allowing the use of longer cables for the link. At data rates higher than 8Gbps, the receive PLL fails due to increased high-frequency noise in the loop. Raising the supply to 3.3V allows the receiver to perform up to 9Gbps. The decision logic of the data phase detector inject undesired charge onto the VCO control line, causing error in sampling clock phases and data detection. This problem is corrected in the new revision of the chip. At 8Gbps over 10 meters, the time window for error-free detection is 50ps. This window is measured by connecting the receive and transmit PLLs to two different clock sources and varying the delay of one clock source until an error is detected.

The receiver data-recovery PLL requires that the input symbols have a minimum peak-to-peak swing of 800mV differential (400mV swing on each line) to acquire lock, and 600mV differential swing to maintain lock. This PLL has a capture range of >20MHz for a symbol stream with one transition per cycle (5 symbols). The frequency acquisition circuit switches the loop control to the data phase detector when there is less than 100kHz frequency difference between the transmitter and receiver reference clocks. The receive PLL has a jitter of 28ps (p-p) and 4ps (rms), when locked to the incoming data signal.

The chip occupies 2mm x 2mm of die area. The transceiver die photo is shown in Fig. 7.

Table 1 Performance Summary

Transmitter performance	
Maximum transmitter rate	8Gbps @ 3V, 10Gb/s @ 3.3V
Output jitter @ 8Gbps	11ps (p-p), 2ps (rms)
Max. eye opening @ 8Gbps	350mV, 110ps (10-m cable)
Max. eye opening @ 10Gbps	200mV, 90ps (10-m cable)
Receiver performance	
Maximum receive rate	8Gbps @ 3V, 9Gbps @ 3.3V
Data PLL jitter @ 8Gbps	28ps (p-p), 4ps (rms)
Data PLL capture range	>20MHz
Min. swing to capture lock	800mV(p-p) differential
Min. swing to maintain lock	600mV(p-p) differential
Data PLL dynamics	BW >30MHz, Ph.m. >48°
Power dissipation @ 8Gbps, 3V	
Output driver	220mW
Analog (2 PLL)	750mW
Input samplers and logic	130mW
Total	1100mW

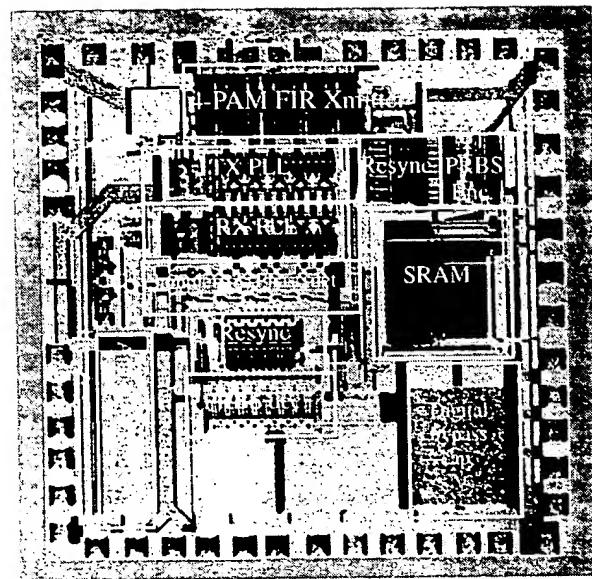


Fig. 7 The transceiver die photograph

Conclusions

Using parallelism, 4-PAM modulation, and analog transmit and receive FIR filters, makes data rates of 8Gbps achievable in conventional CMOS technology over long copper cables. Performance is further enhanced by a novel high-bandwidth linear data-recovery PLL with zero systematic offset that reduces the bit error rate due to random phase errors. A new frequency detector design guarantees frequency acquisition of the data-recovery PLL under all process variations.

Acknowledgments

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